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Response Linearization of a Diode Detector Type Radio Frequency Electric Field Probe

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Response Linearization of a Diode Detector Type Radio Frequency Electric Field Probe

JOSEPH S. ALI, MEMBER, IEEE, GEORGE E. ANDERSON, MEMBER, IEEE, AND E. BAKER BAILEY

Abstract—**An EPROM-based linearization circuit with a resolution of 0.1 percent of full scale has been designed to linearize the response of an orthogonal dipole electric field probe terminated with diodes. Design approach, performance, and probe characteristics are discussed. The nonlinear transducer response is first modeled with a fourth-order polynomial which fits the log-transformed calibration data. The equation is then evaluated at discrete points determined by the memory size and the A/D converter resolution. The antilog of the dependent variable is loaded into an EPROM memory for subsequent readout with a D/A converter and summing network. The response of other nonlinear transducers can be linearized by this method.**

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J. S. Ali is with the Biological Engineering Branch, Experimental Biology Division, Health Effects Research Laboratory, U.S. Environmental Protection Agency, Research Triangle Park, NC 27711.

G. E. Anderson is with the Systems Engineering Branch, Neurotoxicology Division, Health Effects Research Laboratory, U.S. Environmental Protection Agency, Research Triangle Park, NC 27711.

E. B. Bailey is with Northrop Services, Inc., Research Triangle Park, NC 27711.

INTRODUCTION

SMALL non-field-perturbing field probes have long been needed to conduct bioeffects research of radio-frequency radiation. The field probes are used to establish a known exposure power density, map field uniformity, determine the field perturbation caused by animal restrainers, and to measure the field strength inside biological specimens. Most of the field probes commercially available have been designed for surveying applications and, as such, are typically large, have metallic cable conductors, and, in general, are not well suited to the applications listed above. Faced with these circumstances, we constructed small electric field probes using dipole elements terminated with a semiconductor diode. When ex-

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posed to increasing RF field levels, the diode response leaves the square-law region and approaches the linear region of operation causing the detected dc voltage to be a nonlinear function of both electric field strength and power density. When only a single dipole is used, a calibration curve is not too inconvenient. However, more recent field probe designs have incorporated three orthogonal dipoles. The inconvenience of using a distinct calibration curve for each dipole element led us to consider the design of a linearization circuit to simplify the readout of the total equivalent plane-wave power density of the RF field.

Linearization Techniques

Transducer response linearization techniques fall into either of two broad categories—discrete or continuous. Included in the continuous techniques are smooth correction methods and piecewise-linear approximations. Smooth correction methods of linearization have been described using a variety of nonlinear modules such as analog multipliers, dividers, log and antilog amplifiers, and multifunction modules. (The multifunction module raises the input variable to a programmable power as well as implementing multiplication and division operations.) For transducers whose transfer function can be described as a polynomial with a few terms

$$T = aV + bV^2 + cV^3 + \cdots$$

where T and V are the desired and measured values, respectively, a direct power series synthesis of the function can be implemented using multipliers and a summing amplifier [1]. Others [2] have written the transducer response as follows:

$$V = aT + bT^2 + cT^3 + \cdots$$

and solved the equation for T implicitly with the use of multipliers in the feedback loop of the summing amplifier. Good accuracy can be obtained with either method; however, the second method is unstable for certain input voltages and latchup can occur if precautions are not taken.

Visewanath [3] has described a programmable linearization circuit for functions that can be described as the power (not necessarily integral) of some variable. A multifunction module is used to raise the transducer response to the negative of the power characterizing the transducer transfer function thus linearizing the response. Log and antilog amplifiers are also effective in transducer response linearization [4], [5] when the nonlinearity has these functional relationships.

A few methods have been reported using a modified dual-slope A/D converter to linearize as well as digitize the transducer output. Pantic [6] modifies the count accumulated by the A/D converter by blocking one pulse per "correction cycle" where a correction cycle is the number of counts that are allowed to pass to the counters before a single pulse is blocked. The correction cycle is varied by a logic array that monitors the level of the transducer output. This method offers drift-free linearization; however, the linearized output is not monotonic. Weir [7] described a method for modifying an A/D count by using integrators. Ordinarily in a dual-slope A/D converter, the signal is first integrated for a fixed time and then a dc reference of opposite polarity is integrated until a zero crossing is detected. During the discharge cycle, clock pulses are counted. At the zero crossing, the accumulated count is proportional to the signal input. Weir modified the ordinary linear discharge cycle of the integrator by summing to the input of this integrator the output of a second integrator. This method transforms the relationship

$$V = aT - bT^2$$

where T is the desired quantity to

$$V = at - bt^2$$

where t is the integrator discharge time. Since the discharge time is proportional to the number of counts accumulated, the reading is, therefore, proportional to T, the desired variable.

Fig. 1. Orthogonal dipole electric field probe.

Discrete linearization techniques can be implemented either with processor or non-processor instruments. With the former, the transducer nonlinearity is characterized by some analytic function and the function is evaluated at discrete points determined by the resolution of the input A/D converter. For non-processor instruments, an A/D converter can be used to address a ROM which has been programmed with a "lookup table." With the cost of memories plummeting, larger memories can economically be used and with the addition of range-changing circuits before the A/D converter, quantization error can be minimized while using low cost 12-bit A/D converters. The choice between the two methods is generally determined by whether a processor is already in the instrument for some other purpose and whether the typically long function evaluation time is compatible with overall system requirements. Discrete methods have a high degree of stability and do not involve the use of nonlinear components which have higher temperature sensitivity and less accuracy.

In our application, a processor was not needed. This coupled with the advantage of high stability and the need for only one transducer specific adjustment, gain, led us to choose an EPROM linearization method. Even the gain adjustment can be eliminated, if desired, at the expense of some loss in dynamic range since the A/D converter would not be driven to full scale.

RF Field Probe

RF field probe design has been an active area of research in recent years with the most successful and enduring designs having centered around the use of thin-film thermocouple sensors [8], [9] or crystal detectors [10], [11]. The designs, however, were either too large for our applications or were unavailable commercially. The probe that we constructed, shown in Fig. 1, is an isotropic electric field probe using three

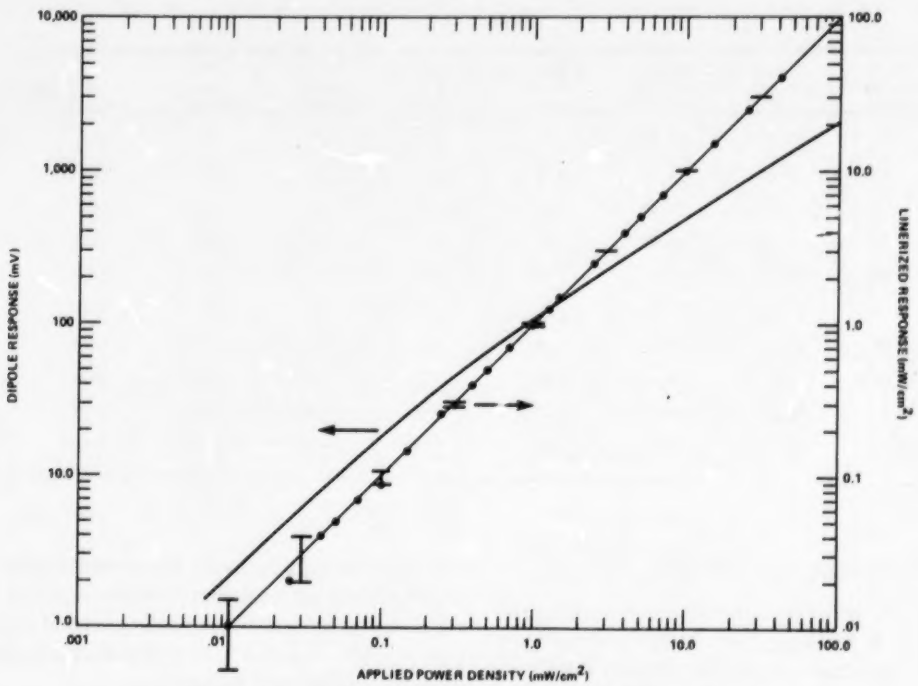

Fig. 2. The curved line is a plot of the Z-channel dipole response characteristic in millivolts (left scale) to an applied power density. The straight line is the desired linearized response (right scale) to an applied power density while the data points on or near this line represent the actual linearizer panel meter readings. The error bars on the linearized response curve show the net quantization error as computed in Table II.

electrically small orthogonal dipoles to produce isotropic response. The design of the probe is an extension of an NBS probe design reported by Swicord [12]. This type of probe design is useful because it can be constructed without specialized equipment and its small size, 3-cm diameter from tip to tip, makes it suitable for high-spacial-resolution field mapping and for measuring equivalent plane-wave power densities in confined areas.

Each dipole element is terminated with a Schottky-barrier diode (HP 5082-2800) at the gap between the wing elements. When exposed to a continuous wave field, a dc voltage is induced across the diode by the rectification of the RF voltage. The voltage is routed to a readout device using NBS developed resistive lines instead of metallic lines to minimize scattering and to reduce coupling to the lines near the antenna elements. The resistive lines used are of two types. Within about 15 cm of the dipoles, a very-high-impedance line of about 70 kΩ/cm is used. A silver epoxy junction joins this line to a 5-ft-long lower impedance line of about 660 Ω/cm. A small shielded cable is then used from this point on to carry the signal to the readout instrument.

The probe tip is constructed using an aluminum mold into which have been bored the outlines of the three dipoles, a spherical cavity, a cavity for the probe handle, and a filling port. The wing elements are $\frac{1}{16}$ in OD stainless steel tubes 1 cm long. The diode lead and a resistive line is slipped into the stainless steel wing element and bonded with silver epoxy. Each dipole is then placed in the mold along with the probe handle and the cavity is filled with polyester resin. Each dipole is individually calibrated from 10 μW/cm^2 to 100 mW/cm^2 at a single frequency between 10 MHz and 1 GHz. A transverse electromagnetic (TEM) transmission cell is used for calibrations between 10 and 500 MHz whereas an anechoic chamber is more suitable from 500 MHz to 1 GHz. Above 1 GHz, the dipole response increases as half-wave resonance is approached.

Although the dipoles are relatively small, above a certain electric field strength the diode response increases beyond the square-law region and enters the transition region. This is shown for the Z-channel dipole response characteristic in Fig. 2. In our application, the equivalent plane-wave power density is the parameter of interest; however, as shown in Fig. 2, the probe will not respond linearly to power density above the square-law region. For a certain size dipole, therefore, either a calibration curve must be used or a linearization technique must be employed.

Detector Response Model

The transducer calibration data have a large dynamic range—the power density has a four-decade range of values while the voltage output varies over a three-decade range. A logarithmic transformation was performed on the data to avoid roundoff errors and to reduce the number of terms in a polynomial curve fit. A model of the form

$$P^1 = B_0 + B_1V^1 + B_2V^{1_2} + B_3V^{1_3} + B_4V^{1_4}$$

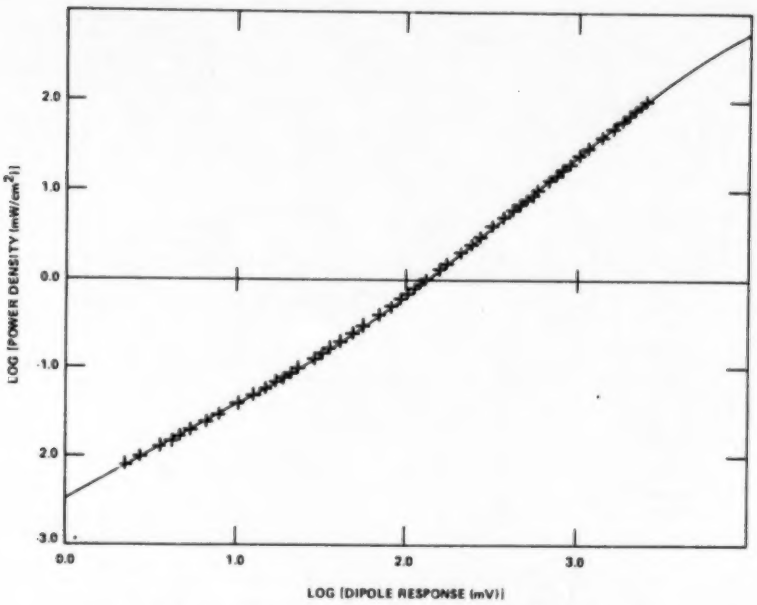

Fig. 3. The data points are the log-transformed Z-channel dipole calibration data while the curved line is a computer-generated fourth-order polynominal curve fit.

where

$$P^1 = \log_{10} [\text{power density in mW/cm}^2]$$

and

$$V^1 = \log_{10} [\text{probe output in mV}]$$

was fitted using a polynomial regression program available for use on the Hewlett-Packard Model 9830A calculator. A plot of the data points and the fitted curve is shown in Fig. 3 for the Z-channel dipole element. Inspection of the residuals demonstrated the appropriateness of the model since the spread of the residuals was uniform over the range of the abscissa and their magnitudes could be explained by the precision with which we were able to collect the data. The coefficient of multiple determination (R^2) was 1.0000 for each of the three sets of calibration data fit with fourth-order polynomials. The coefficients for the three probe element response models are given in Table I.

TABLE I

	X	Y	Z
B_0	-2.3677	-2.3695	-2.4830
B_1	1.1919	1.2143	1.1587
B_2	-0.2819	-0.3092	-0.2369
B_3	0.1874	0.2006	0.1663
B_4	-0.0277	-0.0296	-0.0244

Programming the Memory

A laboratory minicomputer system has been previously interfaced to an EPROM programmer with the appropriate software input/output drivers already written. This greatly facilitated the task of generating the EPROM/DIODE correction table. A Fortran IV program was developed which evaluated the fourth-order polynomials at 1024 equally spaced points over the zero to full-scale input range desired. The polynomial coefficients are input to the program at execution time, allowing the testing and evaluation of various diodes. An array of these power density values (four-digit binary-coded decimal (BCD) notation) is generated so that one EPROM can be programmed with the two most significant digits and the other with the two least significant digits. The program downloads half of the array at a time in Intel hexidecimal object code format to the programmer. A listing of all parameters used and generated is also provided for verification during testing.

Linearizer/Readout Operation

The signal from the dipole element is delivered to a FET input operational amplifier connected in a noninverting amplifier configuration, as shown in Fig. 4. The important factors in selecting this amplifier are low initial input bias current and bias current drift because of the high source impedance caused by the resistive lines, and low initial input offset voltage and offset-voltage drift. For the amplifier used, the initial input bias current and offset voltage are small enough that the amplifier can be operated without an offset-trimming potentiometer. The amplified signal is then applied to a 12-bit binary-coded monolithic CMOS A/D converter. Conversion time is slow (24 ms) because the dual-slope integration method is used, however it is quite adequate for our application and offers the benefits of high accuracy at low cost. Although the A/D converter has

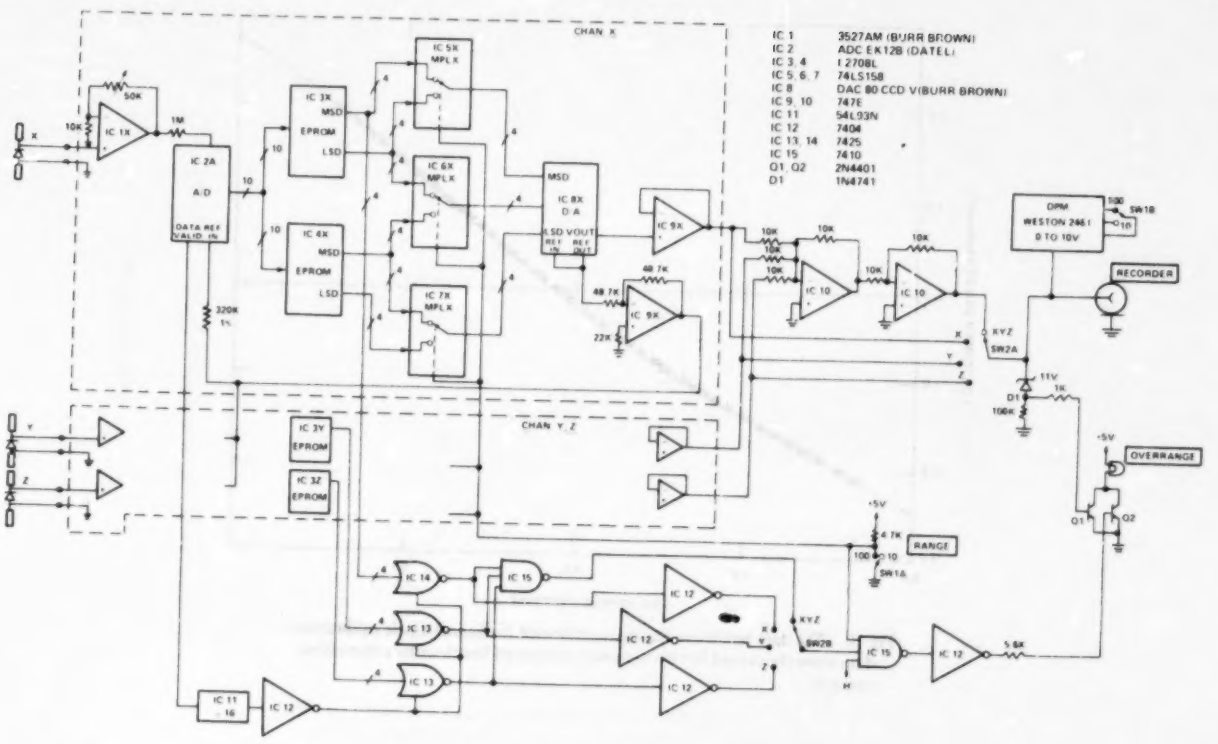

Fig. 4. Schematic drawing of the linearizer electronics.

12-bit resolution, only 10 bits are needed to address the EPROM memory. Each channel has two 2708L EPROMS for storing the linearized calibration curve. The 2708L is configured as 1024 8-bit words and as such each chip can hold two BCD digits. Two chips are addressed in parallel by the A/D converter to yield four BCD digit resolution on the output. This arrangement yields a memory output range of 10 000 to 1. Larger range is needed on the output because of the greater dynamic range of the output variable (power density) compared to the input variable (dipole detector voltage) due to the detector nonlinearity. High resolution is needed only at the lower end of the range, since at the upper end of the range accuracy will be limited by A/D converter quantization error. To avoid the expense of going to a 16-bit D/A converter, a dual-range function with a 12-bit D/A converter was chosen. The full-scale level of the upper range is 99.9 mW/cm^2 while the lower range has a full-scale level of 9.99 mW/cm^2 giving us the desired 10-μW/cm^2 resolution where it is most useful. Range change is accomplished using three quad two-input data selectors (74LS158) per channel. Changing ranges in this way necessitated the use of BCD words in the EPROM memory and BCD input D/A converters. The D/A converter is an industry-standard DAC 80 which is a hybrid integrated circuit with an internal Zener reference. The D/A converter internal reference was also used as a reference for the A/D converter after inversion with IC 9X.

The output of each D/A converter is summed in IC 10 and inverted. Switch 2A selects whether an individual channel or the sum will be displayed. The linearized response for the Z channel is shown in Fig. 2. The signal displayed is also routed to the recorder output jack and, by virtue of the calibration, the voltage is in units of milliwatts per square centimeter.

Overrange Circuitry

Two overrange conditions are sensed by the circuitry. Components D1 and Q1 monitor the output of IC 10 when switch 2A has selected the summing mode. If the output of IC 10 is approaching saturation, D1 and Q1 conduct and the overrange lamp will be lit with a steady glow. The range switch, switch 1A, activates a second overrange circuit only when the low (10-mW/cm^2) range is selected. Recall that the EPROM's in each channel store four decimal digits with a full-scale range of 99.99 mW/cm^2. When in the 10-mW/cm^2 range, the data selectors route only the three least significant digits to the D/A converters. Therefore, the most significant digit must be monitored and an overrange condition indicated if a nonzero condition exists in this digit. The four-input NOR gates (IC 13, 14) provide this monitoring function. In our application we are only concerned if an overrange condition occurs in a selected channel. When switch 2B is in the X, Y, or Z position, only the selected channel can activate the overrange lamp. When the sum of all three channels is being displayed and switch 2B is, therefore, in the XYZ position, then an overrange in any channel will activate the overrange lamp. IC 11 and 12 provide a square wave to the strobe input lines of the NOR gates causing the overrange lamp to flash off and on. This allows the user to distinguish between the two types of possible overrange conditions.

TABLE II

Power Density (mW/cm^2)	Slope ($\frac{\mu W/cm^2}{mV}$)	A/D Conv. Q.E. (RIO) ($\mu W/cm^2$)	Round-Off Error ($\mu W/cm^2$)	DAC Q.E. ($\mu W/cm^2$)	Trunc. Error ($\mu W/cm^2$)	NET Q.E. ($\mu W/cm^2$)	NET Q.E. PWR. DEN. (%)
.01	3.76	±4.66	±0	±5	±0	±5	±50
.03	4.03	±5.00	±10	±5	±0	±10	±33
.10	5.12	±6.35	±10	±5	±0	±10	±10
.30	7.18	±8.90	±10	±5	±0	±10	±3.3
1.00	11.17	±13.9	±10	±5	±0	±13.9	±1.4
3.00	17.10	±21.2	±20	±5	±0	±21.2	±.71
10.00^-	27.5	±34.1	±30	±5	±0	±34.1	±.34
10.00^+	27.5	±34.1	±30	±50	±100	±100	±1.0
30.0	40.9	±50.7	±50	±50	±100	±100	±.33
100.0	58.2	±72.2	±70	±50	±100	±100	±.10

Quantization Error

In a digital-memory-based linearization technique, quantitization error (QE) is developed both by limitations in the number of memory locations as well as the memory word size. Because of the nonlinear relationship between the input to the linearizer (millivolts) and the desired response (milliwatts per square centimeter), the QE is a function of the operating point. The QE at the A/D converter, assuming the use of a 1K × 16 memory is ± one-half the full-scale input voltage divided by 1024. Channel Z, for example, has a full-scale output of 2536.84 mV which yields a QE of ±1.24 mV. Table II shows how this error reflects itself through the sensor transfer function as well as the different contributions that affect the net QE.

The A/D converter QE can be referred to the output (RTO) by taking the product of the A/D QE and the slope of the transfer function. Roundoff errors occur because the lookup table contains corrections stored with four-digit resolution where the least significant digit has units of tens of microwatts per square centimeter. The DAC QE is plus or minus one-half the DAC resolution referred to its input. Remembering that a range change is used at 10 mW/cm^2 and above and that the DAC responds to a 12-bit, 3-digit BCD code, we have that the DAC QE is ±5 $\mu W/cm^2$ below 10 mW/cm^2 and ±50 $\mu W/cm^2$ above 10 mW/cm^2. Finally, truncation error is developed only on the high range when the last digit stored in EPROM is chopped. The net QE is the largest of all the above errors and is shown in Table II both in absolute value and as a percentage of the operating point. The net QE is also shown in Fig. 2 as error bars on the linearized response curve. Above 1.0 mW/cm^2, the net QE remains less than or equal to 1.0 percent becoming only 0.1 percent of reading at full scale.

Conclusion

A digital EPROM-based linearizer has been designed which conveniently displays total equivalent plane-wave power density when used with the herein described diode detector orthogonal electric field probe. The use of a digital range-changing circuit has permitted the use of a low-cost three-digit D/A converter while preserving adequate resolution and acceptable quantization errors. For more stringent linearization requirements, the resolution could be improved by increasing the memory size and implementing several range changes in memory. The linearizer is easy to calibrate requiring only one transducer dependent adjustment.[1]

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[1] The authors have learned through the review process for this paper that a ROM-based linearization technique has also been used by Holaday Industries, Eden Prairie, MN, in the design of some of their RF field strength meters. The authors are unaware of any publication in the open literature describing this work.